REDUCED COMPLEXITY MIMO TURBO EQUALIZATION FOR STTRC-CODES

Nenad Veselinovic, Tad Matsumoto
Centre for Wireless Communications, P.O. Box 4500, 90014 Univ. of Oulu, Finland
tel: +358 8 5532864, fax: +358 8 5532845
e-mail: {nenad.veselinovic,tadashi.mastumoto}@ee.oulu.£

***Abstract*— Turbo MIMO equalization in space-time-trellis-coded (STTrC) system is considered in this paper. First, a low complexity MMSE equalizer is derived for decoding of STTrC-coded signals in frequency selective channels. The receiver is further modi£ed so that only the signi£cant portion of the channel impulse response is taken into consideration. The remaining (non-signi£cant) portion of the channel impulse response is regarded as unknown co-channel interference (UCCI) and suppressed using covariance estimation technique. Its performance is evaluated using measurement data obtained by the multidimensional channel sounder. The receiver performance is shown to be within 2 dB from the maximum-likelihood lower bound, which takes into account all multipath components.**

I. INTRODUCTION

Communications signal transmission and reception using multiple transmit antennas and receive antennas over an multiple-input-multiple-output (MIMO) channel is one of the most promising approaches to increase the link capacity and achievable data rates [1]. Two key approaches have been developed to make effective use of the bene£ts of the MIMO channels. The £rst one is Bell-Labs-Layered-Space-Time-Architecture (BLAST) [2] which aims at approaching the channel outage capacity. Another one that combines the bene£ts of transmit diversity and channel coding is space-time-trellis-coding (STTrC) [3]. Some recent developments combine the bene£ts of the above two approaches [4].

To fully exploit the bene£ts of the broadband, frequency selective channels using single carrier communications, the cost ef£cient implementation of the equalization part of the receiver is a key issue. Furthermore, to fully exploit the capacity of the multipath channel turbo processing has been proposed [5], which turns the multipath channel into a set of parallel diversity channels. Recently, the MMSE-based turbo-equalization has attracted considerable attention due to the possibilities for adaptive implementation [6] and even further complexity reductions [7].

Iterative equalization with STTrC-codes has been introduced in [8], where the optimal MAP equalizer has been used. In this paper, we extend the MMSE-based turbo equalization of [6], [9], [10] to detect STTrC-coded signals. The equalizer is further modi£ed, in order to allow for the performance evaluation using realistic channel impulse responses obtained by using multidimensional channel sounder. Only the signi£cant part of the channel impulse response is taken into account in the reciever, while the rest of multipath components are regarded as the unknown co-channel interference (UCCI) and they are cancelled using covariance estimation technique. The preformance of the receiver is evaluated using several snapshots of the realistic channel impulse responses.

The rest of the paper is organized as follows. Section II describes system model. Section III presents the proposed receiver and its special cases for which either one antenna or all transmit antennas are detected simultaneously. Section IV describes the receiver whose performance is used as a lower bound on the proposed receivers' performance. Section V presents numerical results. The paper is concluded in Section VI.

This work was supported by Nokia, Elektrobit, Finnish Air Forces, Instrumentointi, National Technology Agency of Finland (Tekes), Nokia Foundation and Infotech Oulu Graduate School

II. SYSTEM AND RECEIVED SIGNAL MODEL

Figure 1 describes the system model. Each of K users encodes bit information sequence $c_k(i)$, $k = 1, ..., K$, $i = 1, ..., Bk_0$ using a rate k_0/N_T STTrC code, where N_T and B are the number of transmit antennas and frame length in symbols, respectively. The encoded sequences $b_k(i) \in \mathcal{Q}$, $i = 1, ..., BN_T$ are £rst grouped in B blocks of N_T symbols, where $\mathcal{Q} = \{\alpha_1, \dots, \alpha_{2^{k_0}}\}$ denotes the modulation alphabet assumed to be M-phase-shift-keying (M-PSK). However, it is straightforward to extend the receiver derivations to the quadrature-amplitude-modulations (QAM). The coded sequence is then interleaved so that the positions within blocks of length N_T remain unchanged but the positions of the blocks themselves are permuted within frame according to the user-speci£c interleaver pattern. Thereby the rank properties of the STTrC codes are preserved [11]. The interleaved sequences are then headed by the user-speci£c training sequences consisting of TN_T symbols. The entire frame is serial-to-parallel converted, resulting in the sequences $b_k^{(n)}(i)$, $n = 1, .., N_T$, $i = 1, ..., B + T$ and transmitted with N_T transmit antennas trough the frequency selective channel.

After coherent demodulation in the receiver, the signals from each of N_R receive antennas are sampled in time domain to capture the multipath components. Observing the signals from different transmit antennas of different users as the virtual users and arranging them in the vector form similarly as in [10], [9] we form the space-time representation of the received

0-7803-8523-3/04/$20.00 ©2004 IEEE.

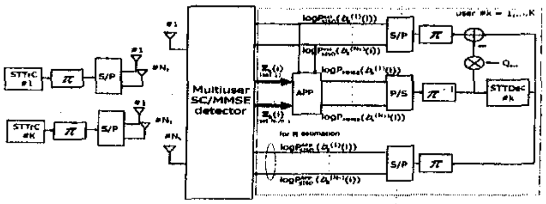

Fig. 1. System model

signal at time instant i given by

$$\mathbf{y}(i) = \underbrace{\mathbf{H}\mathbf{u}(i)}_{\text{desired}} + \underbrace{\mathbf{n}(i)}_{\text{noise}},\ i = 1, ..., T + B, \tag{1}$$

where $\mathbf{y}(i) \in \mathbb{C}^{LN_R \times 1}$ is space-time sampled received signal vector, given by

$$\mathbf{y}(i) = [\mathbf{r}^T(i + L - 1), \ldots, \mathbf{r}^T(i)]^T, \tag{2}$$

with $\mathbf{r}(i) \in \mathbb{C}^{N_R \times 1}$ being

$$\mathbf{r}(i) = [r_1(i), \ldots, r_{N_R(i)}]^T. \tag{3}$$

L is the number of paths of the frequency selective channel and $r_m(i)$ denotes the signal sample obtained after matched £ltering at the mth receive antenna. $\mathbf{H} \in \mathbb{C}^{LN_R \times KN_T(2L-1)}$ is channel matrix with the the form of

$$\mathbf{H} = \text{Toeplitz}(\mathbf{H}(0) \ldots \mathbf{H}(L-1))$$

and

$$\mathbf{H}(l) = \begin{bmatrix} h_{1,1}^{(1)}(l) & \ldots & h_{1,1}^{(N_T)}(l) & \ldots & h_{K,1}^{(N_T)}(l) \\ \vdots & \ddots & \vdots & \ddots & \vdots \\ h_{1,N_R}^{(1)}(l) & \ldots & h_{1,N_R}^{(N_T)}(l) & \ldots & h_{K,N_R}^{(N_T)}(l) \end{bmatrix},$$

where $h_{k,m}^{(n)}(l)$ denotes the l-th path complex gain between kth user's nth transmit antenna and mth receive antenna.

The vector $\mathbf{u}(i) \in \mathbb{Q}^{KN_T(2L-1)\times 1}$ denotes desired users' sequence, and it is de£ned as

$$\mathbf{u}(i) = [\mathbf{b}^T(i + L - 1), \ldots, \mathbf{b}^T(i), \ldots, \mathbf{b}^T(i - L + 1)]^T, \tag{4}$$

with

$$\mathbf{b}(i) = [b_1^{(1)}(i), \ldots, b_1^{(N_T)}(i), \ldots, b_K^{(1)}(i), \ldots, b_K^{(N_T)}(i)]^T, \tag{5}$$

and $\mathbf{n}(i) \in \mathbb{C}^{LM \times 1}$ is the vector containing additive white Gaussian noise (AWGN) with covariance $E\{\mathbf{n}(i)\mathbf{n}^H(i)\} = \sigma^2\mathbf{I}$.

III. Turbo MIMO Equalizers

The receiver £rst associates the the signals from transmit antennas of the kth user to the groups of length n_0, so that antennas indexed by $n = 1, ..., n_0$ belong to the £rst group, those indexed by $n = n_0 + 1, ..., 2n_0$ belong to the second group etc. Obviously, the number of transmit antennas N_T is assumed to be an integer multiple of n_0. Without loss of generality the receiver derivation is presented in Section III-A for the 1st group of transmit antennas of the kth user. The derivation is exactly the same for the rest of transmit antenna groups and the rest of users, with difference only in indexing. The special cases of $n_0 = 1$ (rec. #1) and $n_0 = N_T$ (rec. #2) are considered in more detail in numerical examples.

A. SC/MMSE Equalizer Derivation

Let us rewrite the Eq. 1 so as to take into account only a part of the channel impulse response of length L_{eff}

$$\mathbf{y}_{eff}(i) = \underbrace{\mathbf{H}_{eff}\mathbf{u}_{eff}(i)}_{\text{desired}} + \underbrace{\mathbf{H}_{I1}\mathbf{u}_{I1}(i) + \mathbf{H}_{I2}\mathbf{u}_{I2}(i)}_{\text{interference}} + \underbrace{\mathbf{n}_{eff}(i)}_{\text{noise}}, \tag{6}$$

where $\mathbf{H}_{eff} \in \mathbb{C}^{L_{eff}N_R \times KN_T(2L_{eff}-1)}$, $\mathbf{H}_{I1} \in \mathbb{C}^{PN_R \times KN_T(2P-1)}$ and $\mathbf{H}_{I2} \in \mathbb{C}^{EN_R \times KN_T(2E-1)}$ are de£ned as

$$\mathbf{H}_{eff} = \text{Toeplitz}(\mathbf{H}(P) \ldots \mathbf{H}(L - E - 1))$$

$$\mathbf{H}_{I1} = \text{Toeplitz}(\mathbf{H}(0) \ldots \mathbf{H}(P-1))$$

and

$$\mathbf{H}_{I2} = \text{Toeplitz}(\mathbf{H}(L-E) \ldots \mathbf{H}(L-1))$$

and P, E and L_{eff} are related as $L = P + L_{eff} + E$. The vectors $\mathbf{u}_{eff} \in \mathbb{C}^{KN_T(2L_{eff}-1)\times 1}$, $\mathbf{u}_{I1} \in \mathbb{C}^{KN_T(2P-1)\times 1}$ and $\mathbf{u}_{I2} \in \mathbb{C}^{KN_T(2E-1)\times 1}$ are de£ned as

$$\mathbf{u}_{eff}(i) = [\mathbf{b}^T(i+L_{eff}-1), \ldots, \mathbf{b}^T(i), \ldots, \mathbf{b}^T(i-L_{eff}+1)]^T, \tag{7}$$

$$\mathbf{u}_{I1}(i) = [\mathbf{b}^T(i + L - P - 1), \ldots, \mathbf{b}^T(i+1)]^T, \tag{8}$$

and

$$\mathbf{u}_{I2}(i) = [\mathbf{b}^T(i-1), \ldots, \mathbf{b}^T(i - L - E + 1)]^T. \tag{9}$$

First, an estimate $\hat{\mathbf{H}}_{eff}$ of the channel matrix $\mathbf{H}$ is obtained based on the training sequence $\mathbf{u}_{eff}(i)$, $i = 1, ..., T$ and soft feedback $\bar{\mathbf{u}}_{eff}(i)$ obtained by replacing the corresponding elements of $\mathbf{u}_{eff}(i)$ by their soft estimates, de£ned as

$$\bar{b}_k^{(n)}(i) = \sum_{q=1}^{2^{k_0}} \alpha_q P_{SISO}^{app}(b_k^{(n)}(i) = \alpha_q), \tag{10}$$

where P_{SISO}^{app} denotes *a posteriori* information obtained after SISO decoding. Let us further denote

$$\hat{\mathbf{u}}_k^{(1,n_0)}(i) = \tilde{\mathbf{u}}_{eff}(i) - \tilde{\mathbf{u}}_{eff}(i) \odot \mathbf{e}_k^{(1,n_0)}, \tag{11}$$

where

$$\mathbf{e}_k^{(1,n_0)} = [\ \underbrace{0, ..., 0}_{[(L_{eff}-1)K+k-1]N_T}\ , \underbrace{1, ..., 1}_{n_0},\ \underbrace{0, ..., 0}_{(L_{eff}K-k+1)N_T-n_0}\ ]^T, \tag{12}$$

and $\odot$ denotes elementwise vector product. The vectors $\tilde{\mathbf{u}}_{eff}(i)$ are obtained by replacing the elements of $\mathbf{u}_{eff}(i)$ by their soft estimates, i.e. an element is

$$\tilde{b}_k^{(n)}(i) = \sum_{q=1}^{2^{k_0}} \alpha_q P_{SISO}^{ext}(b_k^{(n)}(i) = \alpha_q), \tag{13}$$

where P_{SISO}^{ext} denotes the extrinsic information obtained after SISO decoding (see [10]). The signals $b_k^{(n)}(i), n = 1, ..., n_0$ are jointly detected by £ltering the signal

$$\hat{\mathbf{y}}_k^{(1,n_0)}(i) = \mathbf{y}_{eff}(i) - \hat{\mathbf{H}}_{eff}\hat{\mathbf{u}}_k^{(1,n_0)}(i), i = T+1, ..., B+T, \quad (14)$$

using a linear MMSE £lter whose weighting matrix $\mathbf{W}_k(i) \in \mathbb{C}^{L_{eff}N_R \times n_0}$ satis£es the following criterion

$$[\mathbf{W}_k(i), \mathbf{A}_k(i)] = \arg\min_{\mathbf{W},\mathbf{A}} ||\mathbf{W}^H\hat{\mathbf{y}}_k^{(1,n_0)}(i) - \mathbf{A}^H\beta_k(i)||^2. \quad (15)$$

The vector $\beta_k(i) \in \mathbb{C}^{n_0 \times 1}$ is de£ned by

$$\beta_k(i) = [b_k^{(1)}(i), ..., b_k^{(n_0)}(i)]^T, \quad (16)$$

and $\mathbf{A}_k(i) \in \mathbb{C}^{n_0 \times n_0}$ is a matrix whose diagonal elements satisfy the constraint

$$a_{1,1}(i) = ... = a_{n_0,n_0}(i) = 1, \quad (17)$$

to avoid the trivial solution $[\mathbf{W}_k(i), \mathbf{A}_k(i)] = [0, 0]$. The columns $\mathbf{w}_k^{(n)}(i) \in \mathbb{C}^{L_{eff}N_R \times 1}$ of the optimal weighting matrix $\mathbf{W}_k(i)$ can be obtained as

$$\mathbf{w}_k^{(n)}(i) = \mathbf{M}_k(i)^{-1}\mathbf{h}_k^{(n)}, \quad (18)$$

where

$$\begin{aligned} \mathbf{M}_k(i) &= \hat{\mathbf{H}}\mathbf{\Lambda}_k(i)\hat{\mathbf{H}}^H + \hat{\mathbf{R}} - \sum_{n=1}^{n_0} \mathbf{h}_k^{(n)}\mathbf{h}_k^{(n)^H}, \quad (19) \\ &= \mathbf{R}_{cov} - \sum_{n=1}^{n_0} \mathbf{h}_k^{(n)}\mathbf{h}_k^{(n)^H}, \end{aligned}$$

and $\mathbf{h}_k^{(n)}$ is the $[(L_{eff} - 1)KN_T + kN_T + n]$-th column of the matrix $\hat{\mathbf{H}}$. The matrix $\mathbf{\Lambda}_k(i)$ is de£ned as

$$\begin{aligned} \mathbf{\Lambda}_k(i) &= \mathbf{I} - E\{\hat{\mathbf{u}}_k^{(1,n_0)}(i)\hat{\mathbf{u}}_k^{(1,n_0)}(i)^H\} \quad (20) \\ &= \mathrm{diag}\{1 - |[\tilde{\mathbf{u}}(i)]_1|^2, ..., 1 - |[\tilde{\mathbf{u}}(i)]_{(L_{eff}-1)KN_T}|^2, \\ &\quad \underbrace{1, ..., 1}_{n_0}, 1 - |[\tilde{\mathbf{u}}(i)]_{(L_{eff}-1)KN_T+n_0+1}|^2, ..., \\ &\quad 1 - |[\tilde{\mathbf{u}}(i)]_{(2L_{eff}-1)KN_T}|^2\}. \end{aligned}$$

It can be shown that the covariance matrix $\mathbf{R}$ of the interference-plus-noise has the following form

$$\begin{aligned} \mathbf{R}(i) &= \hat{\mathbf{H}}_{eff}\mathbf{\Lambda}_{k,1}(i)\mathbf{H}_{I1}^H + \hat{\mathbf{H}}_{eff}\mathbf{\Lambda}_{k,2}(i)\mathbf{H}_{I2}^H \quad (21) \\ &+ \mathbf{H}_{I1}\mathbf{H}_{I1}^H + \mathbf{H}_{I2}\mathbf{H}_{I2}^H + \sigma^2\mathbf{I}. \end{aligned}$$

In this paper we use the time-average approximation as follows

$$\begin{aligned} \hat{\mathbf{R}} &= \frac{1}{T}\sum_{i=1}^{T} ||\mathbf{y}_{eff}(i) - \hat{\mathbf{H}}_{eff}\overline{\mathbf{u}}_{eff}(i)||^2 \quad (22) \\ &+ \frac{1}{B}\sum_{i=T+1}^{T+B} ||\mathbf{y}_{eff}(i) - \hat{\mathbf{H}}_{eff}\overline{\mathbf{u}}_{eff}(i)||^2. \end{aligned}$$

Matrices $\mathbf{\Lambda}_{k,1}(i)$ and $\mathbf{\Lambda}_{k,2}(i)$ are de£ned as $\mathbf{I} - E\{\hat{\mathbf{u}}_k^{(1,n_0)}(i)\tilde{\mathbf{u}}_{I1}(i)^H\}$ and $\mathbf{I} - E\{\hat{\mathbf{u}}_k^{(1,n_0)}(i)\tilde{\mathbf{u}}_{I2}(i)^H\}$, respectively. Note that Eq. (20) holds only for the M-PSK case, although it is straightforward to extend the receiver derivation to the more general signal constellations. Assuming that the MMSE £lter output $\mathbf{z}_k(i) \in \mathbb{C}^{n_0 \times 1}$ can be viewed as the output of the equivalent Gaussian channel we can write

$$\begin{aligned} \mathbf{z}_k(i) &= \mathbf{W}_k^H(i)\hat{\mathbf{y}}_k^{(1,n_0)}(i) \quad (23) \\ &= \mathbf{H}_{e,k}(i)\beta_k(i) + \mathbf{\Psi}_{e,k}(i), \end{aligned}$$

where matrix $\mathbf{H}_{e,k}(i) \in \mathbb{C}^{n_0 \times n_0}$ contains the channel gains of the equivalent channel de£ned as

$$\mathbf{H}_{e,k}(i) = E\{\mathbf{z}_k(i)\beta_k^H(i)\} = \mathbf{W}_k^H(i)\mathbf{H}_{ML,k}, \quad (24)$$

with $\mathbf{H}_{ML,k} = [\mathbf{h}_k^{(1)} \dots \mathbf{h}_k^{(n_0)}]$. The vector $\mathbf{\Psi}_{e,k}(i) \in \mathbb{C}^{n_0 \times 1}$ is the equivalent additive Gaussian noise with covariance matrix

$$\begin{aligned} \mathbf{R}_{e,k}(i) &= E\{\mathbf{\Psi}_{e,k}(i)\mathbf{\Psi}_{e,k}^H(i)\} \quad (25) \\ &= \mathbf{W}_k^H(i)\mathbf{R}_{cov}\mathbf{W}_k(i) - \mathbf{H}_{e,k}(i)\mathbf{H}_{e,k}(i). \end{aligned}$$

The output of the equivalent channel $\mathbf{z}_k(i)$ and its parameters $\mathbf{H}_{e,k}(i)$ and $\mathbf{R}_{e,k}(i)$ are passed to the APP block that calculates the *extrinsic* probabilities needed for SISO decoding, as described in section III-B. The similar procedure is repeated for all N_T/n_0 groups of transmit antennas that are jointly detected. It should be noted that different values of $\mathbf{z}_k(i)$, $\mathbf{H}_{e,k}(i)$ and $\mathbf{R}_{e,k}(i)$ are obtained for each group and the dependency of these parameters on the group index is omitted from notation for simplicity.

B. APP Block and SISO Decoding

The SISO channel decoding algorithm used in this paper is a symbol-level *maximum-a-posteriori* (MAP) algorithm used in [10]. It should be noted that the input required by the decoder is the probability $P(S_i, S_{i+1})$ associated with the transition between two trellis states S_i and S_{i+1} of the STTrC code. The transition probability can be calculated as

$$\begin{aligned} P(S_i, S_{i+1}) &= P(\beta_k(i) = \mathbf{d}^{i,i+1}) \quad (26) \\ &= \prod_{n=1}^{N_T} P_{MMSE}^{ext}(b_k^{(n)}(i) = d_n^{i,i+1}), \end{aligned}$$

where $\mathbf{d}^{i,i+1} \in \mathbb{C}^{N_T \times 1}$ is the vector of encoder outputs, that are associated with the transition (S_i, S_{i+1}). The probabilities $P_{MMSE}^{ext}(b_k^{(n)}(i) = \alpha_q)$ are extrinsic probabilities obtained by the MMSE detection, which are calculated in the APP block as

$$P_{MMSE}^{ext}(b_k^{(n)}(i) = \alpha_q) = \quad (27)$$
$$\sum_{\mathbf{f} \in \mathbf{B}^{d_n}} P(\mathbf{z}_k(i)|\mathbf{f}) \left[\prod_{p=1, p \neq n}^{n_0} P_{SISO}^{ext}(b_k^{(n)}(i) = d_n)\right],$$

for $q = 1, ..., 2^{k_0}$ and $n = 1, ..., n_0$, where $\mathbf{B}^{d_n} = \{\mathbf{f} \in \mathcal{Q}^{n_0 \times 1} | f_n = d_n\}$ and

$$P(\mathbf{z}_k(i)|\mathbf{f}) = e^{-(\mathbf{z}_k(i) - \mathbf{H}_{k,e}(i)\mathbf{f})^H \mathbf{R}_{e,k}^{-1}(i)(\mathbf{z}_k(i) - \mathbf{H}_{k,e}(i)\mathbf{f})}, \quad (28)$$

Based on the transition probabilities $P(S_i, S_{i+1})$ the SISO channel decoder calculates the *a posteriori* probabilities for the symbols $b_k^{(n)}(i)$, defined as

$$P_{SISO}^{app}(b_k^{(n)}(i) = \alpha_q) = P(b_k^{(n)}(i) = \alpha_q | \mathbf{z}_k(i), \mathbf{H}_{k,e}(i), \mathbf{R}_{e,k}(i), i = T+1, ..., T+B) \quad (29)$$

The decoder extrinsic probability is then calculated as

$$P_{SISO}^{ext}(b_k^{(n)}(i) = \alpha_q) = \frac{P_{SISO}^{app}(b_k^{(n)}(i) = \alpha_q)}{\left[P_{MMSE}^{ext}(b_k^{(n)}(i) = \alpha_q)\right]^{0.5}}. \quad (30)$$

The receiver complexity is dominated by the MMSE part which requires inversion of the matrix $\mathbf{M}_k(i)$ as well as by the APP block. The overall complexity is therefore $O\{\max(L_{eff}^3 N_R^3, 2^{k_o n_o})\}$.

IV. Numerical Examples

Performance of the proposed receivers was evaluated through computer simulations. The 4-state QPSK code with $N_T = 2$ presented in [3] was used to encode signals of all MIMO users. The Log-MAP space-time trellis decoder shown in [12] and [10] was used. The user specific random interleavers were assumed. To evaluate receiver performance in a realistic scenario channel measurement data for micro cell SIMO scenario available in [13] was used in simulations. Figs. 2 a) show a measurement route as well as the RMS spatial spread at the transmitter side along the route. The data is collected along the route ST9-ST12 indicated in Fig. 2 a). Omnidirectional sleeve receive antenna was moved along the measurement route. The transmit antenna was uniform linear array (ULA) with 8 elements. Two snapshot points at positions 2018 and 4038 are selected in our simulations and receive antennas 1 and 8 were used. The corresponding channel impulse responses are presented in Figs. 2 c) and d). The power of each snapshot was normalized so that the average received power over all snapshots is constant and equal to unity.

In Fig. 3 the SER performances of receiver #1 is presented vs. timing offset P for two chosen channel impulse responses. The optimal values $P = 1$ and $P = 4$ are selected correspondingly. In Figs. 3a) and b) the FER vs. E_s/N_0 is presented for both considered receivers ($n_0 = 1$ and $n_0 = 2$). The receiver #2 shows somewhat better performance comparing to the receiver #1 due to the fact that the two transmit antennas are jointly detected. Receivers #2 and #1 perform within 2 and 3dB from the ML lower bound, respectively.

V. Conclusions

A new iterative receiver scheme for the STTrC coded multiuser system in frequency selective channels is derived. The receiver was modifed so that only signifcant portion of the channel impulse response is used, while the less signifcant portion of the channel impulse response is regarded as the unknown cochannel interference. It was verifed trough computer simulations using realistic channel measurement data that the proposed receiver preforms relatively close to the ML receiver. The receiver that jointly detects signals from all the transmit antennas of the user of interest outperforms the receiver that detects only one antenna at a time. Future work will include receiver performance dependence on the angular spread at the transmitter side.

VI. Acknowledgement

The authors gratefully acknowledge MEDAV for providing the measurement data, Mr. Christian Schneider and Mrs. Mariella Saarestoniemi for providing measurement data conversion software and Prof. Markku Juntti for fruitful discussions and comments.

References

[1] E. Telatar, "Capacity of multi-antenna gaussian channels," *European Trans. Telecommun.*, vol. 10, no. 6, pp. 585–595, Nov.-December 1999.

[2] G.J. Foschini, "Layered space–time architecture for wireless communication in a fading environment when using multi-element antennas," *Bell Labs Tech. J.*, pp. 41–59, Aug. 1996.

[3] V. Tarokh, N. Seshadri, and A. R. Calderbank, "Space-time codes for high data rate wireless communication: Performance criterion and code construction," *IEEE Trans. Inform. Theory*, vol. 44, no. 2, pp. 744–765, 1998.

[4] A. O. Berthet and R. Visoz, "Iteraive decoding of concatenated layered space-time trellis codes on MIMO block fading multipath AWGN channel," *IEEE Trans. Commun.*, vol. 51, no. 6, pp. 940–952, June 2003.

[5] S. L. Ariyavisitakul, "Turbo space-time processing to improve wireless channel capacity," *IEEE Trans. Commun., submitted*, 2000.

[6] X. Wang and H. V. Poor, "Iterative (turbo) soft interference cancellation and decoding for coded CDMA," *IEEE Trans. Commun.*, vol. 47, no. 7, pp. 1046–1061, July 1999.

[7] K. Kansanen and T. Matsumoto, "A computationally effcient MIMO turbo-equaliser," in *Proc. IEEE Veh. Technol. Conf. (VTC)*, Jeju, Korea, Apr. 22–25 2003, vol. 1.

[8] G. Bauch, A. F. Naguib, and N. Seshadri, "MAP equalization of space-time coded signals over frequency selective channels," in *Proc. IEEE Wireless Commun. and Networking Conf. (WCNC)*, New Orleans, USA, Sept. 1999, pp. 261–265.

[9] T. Abe and T. Matsumoto, "Space-time turbo equalization in frequency-selective MIMO channels," *IEEE Trans. Veh. Technol.*, vol. 52, no. 3, pp. 469–475, May 2003.

[10] B. Lu and X. Wang, "Space-time code design in OFDM systems," in *Proc. IEEE Global Telecommun. Conf. (GLOBECOM)*, San Francisco, CA, USA, November 27 – December 1 2000, vol. 2, pp. 1000–1004.

[11] A. F. Naguib, V. Tarokh, N. Seshadri, and A. R. Calderbank, "A space-time coding modem for high-data-rate wireless communications," *IEEE J. Select. Areas Commun.*, vol. 16, no. 8, pp. 1459–1478, 1998.

[12] S. Benedetto, G. Montorsi, D. Divsalar, and F. Pollara, "Soft-in-soft-output APP module for iterative decoding of concatenated codes," *IEEE Commun. Lett.*, vol. 1, no. 1, pp. 22–24, jan 1997.

[13] "www.channelsounder.de," .

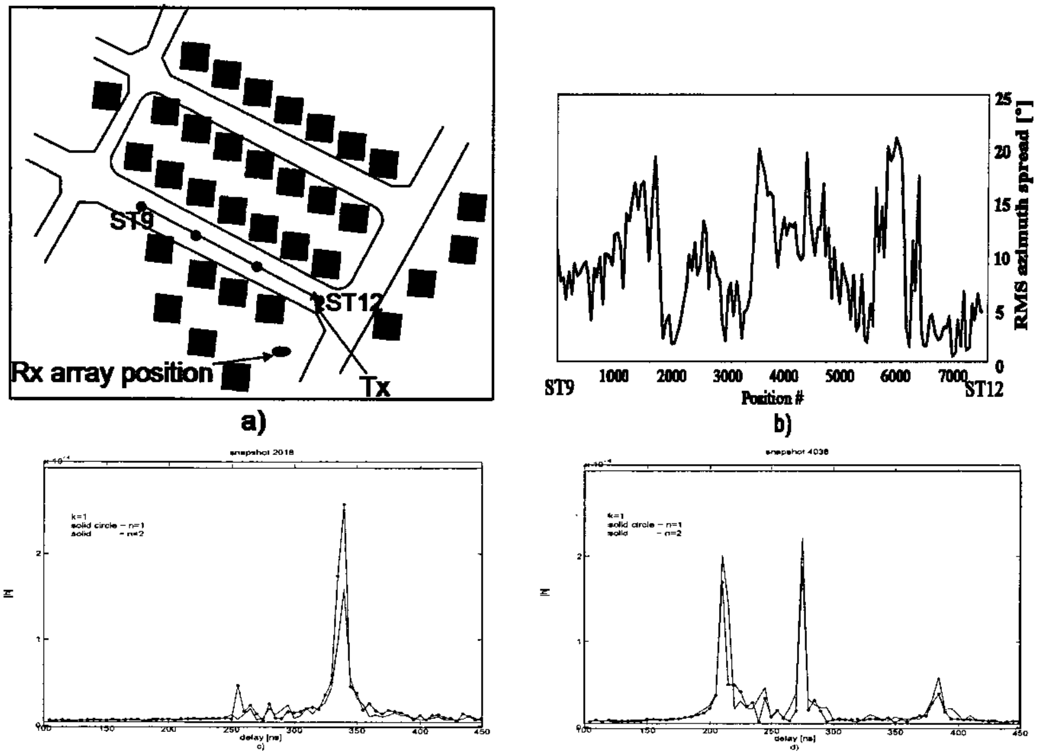

Fig. 2. Measurement scenario, a) map of the area, b) Rx angular spread, c) channel impulse response from antennas 1 and 8 for position 2018, d) channel impulse response from antennas 1 and 8 for position 4038

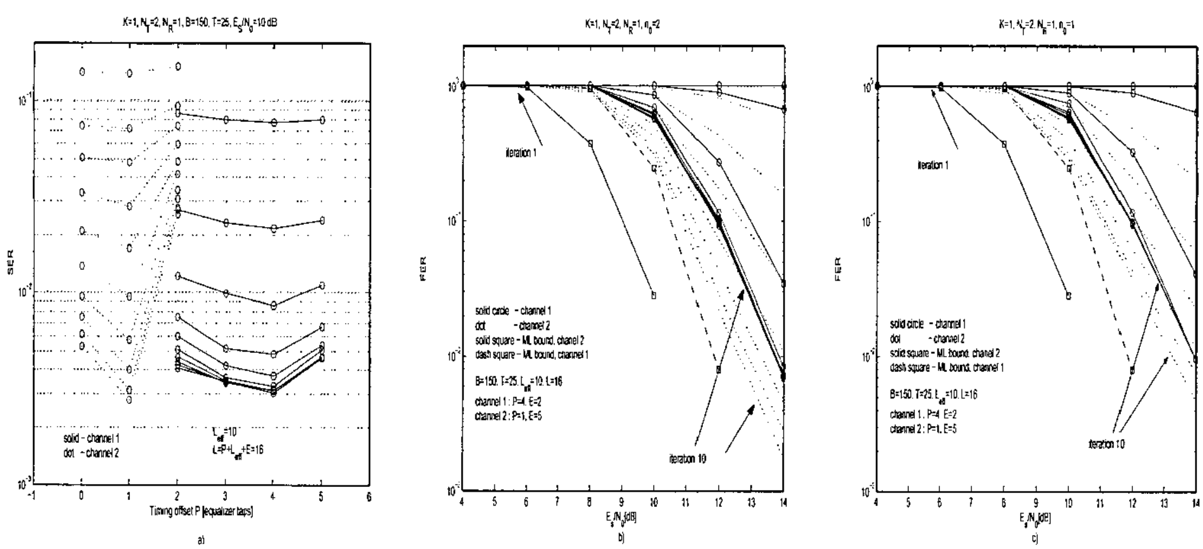

Fig. 3. a)SER performance vs. timing offset P, b) FER performance for $n_0 = 2$, b) FER performance for $n_0 = 1$